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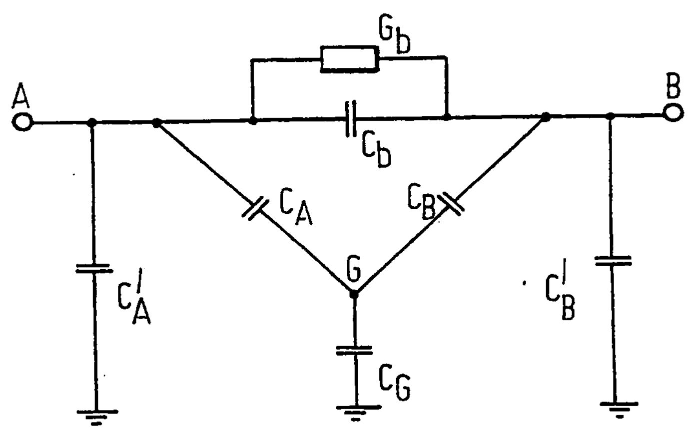

(57) Abstract

A process and an instrument for determining the proportions of gas, water and oil in a gas/water/oil mixture. In the process the permittivity and the dielectric loss in the three-component mixture is being measured during passing said mixture in a flow through a spacing between two opposite electrodes. Signals received from said electrodes are processed in a computer by incorporating specific equations therein and progressively determining said proportions. The instrument comprises an electrode impedance sensor located in a concentrated flow of said three-component mixture, an impedance sensor head, a screen, impedance electronics comprising a tuned amplifier and a phase locked loop circuit forming a tracking resonance system, and a compute for processing the electronic signals and determining said proportions.

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THREE COMPONENT RATIO MEASURING PROCESS AND THREE COMPONENT RATIO MEASURING INSTRUMENT

This invention relates to a process and an instrument for determining the proportions of the components of a three-component mixture and in particular to a process and an instrument for determining the proportions of gas, water and oil in a gas/water/oil mixture.

This invention is applicable on different types of three-phase mixtures, but will herein be described mainly with reference to a three-phase mixture containing gas (g), water (w) and oil (o). Accordingly, the equations stated herein are referring to a three-phase mixture containing gas, water and oil and to the three specific components thereof. However, the equations stated herein are generic to different components, i.e. in general to three components stated as components g, w, and o, respectively.

In crude oil production, the fluid at the well head rarely, if ever, involves single phase single component flow. Usually it will contain crude oil, gas (in the free state and/or dissolved in the oil) and possibly water with the proportion of the latter tending to increase as production continues.

Measurements of the rates of hydrocarbon production from the wells in a field are required for reservoir management and production allocation. With such data, depletion of the reservoir can be undertaken with the aim of optimising total production over the field life. In addition the metering of the total production rate of oil and gas from a field is required for fiscal purposes. This demands a much higher degree of accuracy.

At present, crude oil production systems require separation of the gas, water and oil phases before satisfactory metering, for whatever reason, can be achieved.

Offshore production platforms of today, e.g. those in the North Sea, have a manifold system which allows the flow from a production well to be directed either to the main separators or to a test separator.

Thus at any time, although usually to a set cycle, the flow from a single well can be directed to a test separator where it is separated into gas, water and oil. Meanwhile the production from all other wells is manifolded together and processed in the main separation system. The flowrates of gas, water and oil are measured in the respective single phase lines from the test separator. All three flows are normally measured by using orifice plates, turbine meters or other conventional equipment. A considerable proportion of remaining oil reserves is believed to lie offshore under water depths in excess of 200 metres, in relatively small oil fields, and in hostile environments. As any one of these conditions intensifies, and more particularly when two or more are present together, the cost of conventional offshore recovery systems wherein drilling and production facilities are mounted on surface platforms rises rapidly and soon becomes uneconomic.

For this reason attention has been given to subsea systems where a favoured technique is to drill a number of wells close together and to mount the well head control equipment on the sea bed.

In any design for a new production facility the need for metering of the flows from individual wells, and of the bulk output of the field, must be considered in detail. With the advent of alternative production system concepts such as those outlined above it has become apparent that new methods of flow measurement are desirable to improve both the technical and economic viability of such projects.

One particular problem with multi-phase regimes is the variability of flow conditions in the line. Stratified flow, wavy flow, bubble flow, plug flow, slug flow and annular flow can all occur at various times in horisontal lines. Vertical flow avoids stratification, however.

Dykesteen et al, J. Phys. E; Sci Instrum., Vol, 18, 1985, 540-544 proposes a method for the non-intrusive measurement of th component ratio of a gas/water/oil mixture. The fraction of gas, water and oil flowing between two insulated electrodes is determined by measuring both the resistance and capacitance across the sensor. A mathematical model is used to relate these measured values to the void fraction and water fraction of the flow.

We have now devised an instrument for three component ratic measurement based on the impedance principle. The electrical impedance is measured across two electrodes between which an gas/water/oil mixture is flowing. The impedance is dependent on the geometry and construction of the sensor, the dielectric properties of the individual fluids, the flow regime, the void fraction α and the water fraction β and the oil fraction γ of the flow. However, if the first three factors are constant, then the measured impedance will be a direct function of the component ratio.

According to the present invention there is provided a process for determining the proportions of the components of a three-component mixture, especially the proportions of gas, water and oil in a gas/water/oil mixture. The process of the invention is characterised by measuring the permittivity and the dielectric loss in the three-component mixture during passing said mixture in a flow through a spacing between two opposite electrodes, by processing signals received from said electrodes in a computer by incorporating specific equations therein, and by progressively determining the proportions of said three components in said three-component mixture.

It is thus possible to determine in succession and with adequate accuracy the three components of the mixture flow at rather short intervalls during the passage of the mixture flow in the spacing between the electrodes.

Further, according to the present invention there is provided an instrument for determining the proportions of the components in a three-component mixture, and especially the proportions of gas, water and oil in a gas/water/oil mixture. The instrument is characterised by an electrode impedance sensor located in a flowline of said three-component mixture for

measuring the permittivity and the dielectric loss of said three-component mixture, an impedance sensor head for reducing or eliminating impedance between the electrodes and a screen, impedance measurement electronics comprising a tuned amplifier and a phase locked loop circuit forming a tracking resonance system, for maintaining the resonance frequency, and a computer for processing the electronic signals and determining the proportions of said three components (gas, water and oil) in said three-component mixture.

The present invention, i.e. the process as well as the instrument of the invention, has the following advantages:

- non intrusive -
- measure complete flow through pipe, no sampling -
- real time -
- no moving parts -
- reliable with low maintenance.

By incorporation of a gamma densitometer in the instrument a periodic calibration of the instrument is made possible.

<u>The impedance measurement electronics</u>

Two main objects of the present invention are 1) to provide for a high accuracy in resistance measurements and 2) to provide for a rapid measuring system with instantaneous measures in a sensor with an axial extension of say 0.5 m and with a mixture flow at a rate of say 5 m/s.

In this respect the measurements of capacitance do not represent complications in respect of accuracy. In practice, when ordinary occuring mixtures of oil, water and gas is delivered from an crude oil well the capasitive reactance of the mixture tends to be from about 40 to 400 times less than the resistance in the mixture. In order to avoid that the reactive current should completely exceed the resistive current and thus complicate the measurement of the equivalent resistance of the mixture, the application of a tuned system has been choosen. Accordingly, the sensor capacitance is being tuned with a suitable inductance (L = 1mH) and the system is excited automatically at correct frequency through a phase locked loop

circuite. The resonance frequency provide for a direct measure of the sensor capacitance and accordingly, will not require a separate measuring system.

The dominant reactive current will thus circulate between the sensor capacitance and the tuned inductance. The sole current that is to be supplied in the system is the resistive one to substitute the loss of effect in the mixture flow and in the tuning coil and totally this loss of effect amount to about $15 \cdot 10^{-6}$W. This current separately allows measurement of high accuracy.

The choice of operating frequency for the tracking resonance system is a trade-off between several factors. Firstly, the higher operating frequency we choose, the higher the resistive current will be in relation to the system noise level. Accordingly, this reduces the requirements to the tracking resonance system, and hence, from this point of view, a frequency as high as possible is recommended. There is, however, an upper frequency limit given by the band-with of the selected operational amplifier in combination with the variations in the sensor stray capacitances. The choice of operating frequency is futher restricted by available core material of the tuning coil. An upper frequency of say 510 kHz has been found to be a sensible trade-off between the different factors mentioned. The choice of the low frequency limit is, however, less important, as such lower frequency is caused by a high water consentration in the sensor and an associated low loss resistance.

Because of temperature sensitivity, the impedance electronics should be thermostatically controlled. Conveniently the circuit may be placed in a thermally insulated container and maintained at constant temperature, e.g. 47°C in an atmosphere of nitrogen.

The impedance sensor

Situating the sensor inline in the production system means that very little pressure drop can be tolerated across the sensor. Furthermore in a riser, erosion can be very high and therefore the sensor should ideally be non-intrusive.

Taking these factors into account the electrodes of the sensor head are in the shape of arcuate electrode plates flushing with an inner surface of a tube forming sensor head, in the following denoted as a surface plate sensor.

Insulating the electrodes from the mixture flow reduces the sensitivity of the sensor. Accordingly, and by preference, the radially innermost surface of the electrode plates are in direct, i.e. un-insulated contact with the mixture flow.

In the surface plate electrode sensor, the electrical field penetrates through the whole measuring section, making the sensor sensitive to the flow in the centre of the pipe as well as along the pipe wall. Careful choice of the electrode opening angle ensures that every part of the measuring volume will have the same effect on total sensor impedance.

It can be shown that with respect to homogeneity of the electric field, the optimum electrode opening angle is between 60° and 90°. The sensitivity of a surface plate sensor with such an opening angle will also be good.

In accordance with a preferred embodiment of the invention the electrodes, circumferencially along the inner surface of the tube forming sensor head, are mutually spaced by an opening angle of between approximately 60 to 90°.

The sensor head is ring shaped, or sleeve shaped, with an radially outermost house forming screen, a pair of radially innermost electrodes and, in a volume between same, it contains an electrically insulating material.

A guard is located between the screen and at least a first one of the electrodes, and said guard is at the potential of said first electrode. The guard thereby reduces or eliminates impedance between the electrodes and the screen.

The outer surface of the sensor will in practice be in electrical contact with the rest of the pipe, which means that the outer screen accordingly, by preferance, is at earth potential.

Accordingly the sensor head screens or guards the surface electrodes from electrical noise and other unwanted effects.

While the sensor house will normally be made from steel, the electrodes may be made from steel, or other material having a suitable conductivity such as conducting ceramics.

The separation between the electrodes and guard, and between the guard and the sensor housing must be chosen as a compromise between keeping a low capacitive load on the driver circuit, versus a practical overall size of the sensor. As the separation between the different electrodes increases, the capacitive load to be driven by the guard driver decreases. At the same time the overall diameter, and hence, also the weight of the sensor, increases. The necessary separation is also being influenced by the dielectric properties of the insulating material. A low permittivity insulating material allows the separation to be reduced.

Other criteria for choise of insulating material is that it should have low dielectric loss, and low water absorption. One such material fulfilling these criteria is polyurethane. However, in circumstances where erosion might be a problem, e.g. in a production riser, it may be replaced by a wear resistant insulating material such as a ceramic. A ceramic insulator will also improve the termal stability of the sensor, while the permittivity is higher than for polyurethane, causing the overall size and weight of the sensor to increase.

The impedance electronics is used together with the surface plate electrode sensor head in order to measure the dielectric properties of the flowing mixture and to transmit the data to a processing unit.

Impedance Measurement Only

The deviced impedance measurement system can measure both the gas and water fractions by measuring the permittivity and dielectric loss of the three-component mixture. By experience we know that this system measures the water fraction usually with a precision to within 3% of full range, while the precision of the gas fraction measurement may in some cases be worse than 10% of full range. The reason for this is the sensitivity of the gas fraction to errors in the measured dielectric loss of the mixture. Other sources of error for which one can compensate include the temperature and pressure of the flow.

From these two measurements gas fractions α and the water fraction β are calculated (with the oil fraction γ making up the remainder) from the following equations (1) and (2).

The permittivity of a three component mixture can be related to the component volume fractions by the following relationship:

$$\frac{1 - \varepsilon_b}{1 - \varepsilon_s} \left(\frac{\varepsilon_s}{\varepsilon_b}\right)^{1/3} = 1 - \alpha \qquad [1]$$

where α is the gas fraction, and ε_b is permittivity of the three-component mixture. ε_s is the permittivity of a two-component (oil/water component) of the three-component mixture, and is given by:

$$\varepsilon_s = \frac{\varepsilon_o}{(1 - \beta_f)^{3-4.2 \cdot \beta_f^6 + 3 \beta_f^{10}}} \qquad [1a]$$

where β_f is the water concentration in the two-component (oil/water component). This formula is a slight modification of Bruggemans formula for two-component mixtures. The actual water fraction β of the three-component mixture is given by:

$$\beta = (1 - \alpha)\beta_f \qquad [1b]$$

The dielectric loss of a three-component mixture can be related to the component volume fractions by the following relationship

$$\sigma_b = \frac{\sigma_s}{\varepsilon_s} \{\varepsilon_b - \alpha[\alpha + (\frac{3\varepsilon_s}{1+2\varepsilon_s})^2(1-\alpha)]\} \qquad [2]$$

where σ_s is the dielectric loss of the two-component (oil/water) mixture, ε_s is the two-component (oil/water) permittivity, and ε_b is permittivity of the three-component mixture. σ_s is theoretically given by:

$$\sigma_s = \sigma_o \frac{\varepsilon_s}{\varepsilon_o} \qquad [2a]$$

where ε_s is defined above.

Impedance and Density Measurement

The density of a three-component gas/water/oil mixture is given by:

$$\rho_m = \alpha\rho_g + \beta\rho_w + (1-\alpha-\beta)\rho_o \qquad [3]$$

where ρ_m is the density of the three-component mixture, ρ_g is the gas density, ρ_w is the water density and ρ_o is the oil density. This ρ_m is the density that is measured by a gamma densitometer. If we somehow know the water fraction β, we can calculate the gas fraction α from Eq. 4 :

$$\alpha = \frac{\rho_m - \beta(\rho_w - \rho_o) - \rho_o}{(\rho_g - \rho_o} \qquad [4]$$

Conversely if the gas fraction is known, Eq. 4 can be used to compute the water fraction. From this it is obvious that a gamma densitometer alone cannot be used to measure either the gas or water fraction. However, in order to improve the accuracy of the gas measurement, the impedance system can be combined with a gamma densitometer by feeding the measured water fraction from the impedance system into Eq. 3. Assuming that the densities of the three components are known with sufficient precision, the accuracy of α from Eq. 3 is in fact as good as the accuracy of the β from the impedance system. Having obtained a good value for α, this can be put into the three-component permittivity-formula to compute a new value for β that is even more accurate than the original β from the impedance system. This new β can then be put into Eq. 3 to compute another α of still higher precision and so on in any desired number of iterations.

The gamma densitometer output has to be integrated over several seconds to give an accurate result. In some flow regimes the gas and water fractions can vary rapidly, and as the relation between these fractions and the mean permittivity and mean density are non-linear, averaging the two measurements independently, before calculating the fractions, can lead to errors. In this implementation (active reference) the fractions are calculated from the impedance measurement, and the averaged fractions are compared with fractions as calculated in the above described process. Any difference is used to calibrate the impedance measurement circuits, thus improving the (instantaneous) accuracy of the impedance sensor.

The invention is illustrated by, but not limited with reference to Figures 1-16 of the accompanying drawings wherein

Figure 1 is a schematic representation of a system incorporating an impedance sensor used to determine the gas fraction α and the water fraction β, with the oil fraction γ obtained by the difference.

Figure 2 is a schematic representation of a system according to Figure 1 with a gamma-densitometer used as an active reference.

Figure 3 is a section through a sensor.

Figure 4 is a circuit diagram of a sensor head equivalent circuit.

Figure 5 is a circuit diagram of a virtual ground circuit.

Figure 6 is a circuit diagram of a driven sensor circuit.

Figure 7 is a circuit diagram of a driven sensor equivalent circuit.

Figure 8 is a circuit diagram of a simplified equivalent circuit.

Figure 9 is a block diagram of Figure 5.

Figure 10 is a re-arranged block diagram.

Figure 11 is a tuned driven sensor.

Figure 12 is a equivalent circuit of the operational amplifier.

Figure 13 is a plot of resonance frequency vs. bulk capacitance.

Figure 14 is a plot of bulk loss conductance vs. gain and bulk capacitance.

Figure 15 is a block diagram of a tracking resonance system.

Figure 16 is a section trough a sensor in an alternativ embodiment of the one illustrated in Figure 3.

With reference to Figure 1, a three-component mixture of gas, water and oil flows through a line 1 incorporating an impedance sensor 2 which measures the permittivity and dielectric loss of the mixture. A transmitter 3 feeds signals from the sensor 2 to a data processing unit (computer) 4 which can incorporate a display of the relative proportions of the three-component mixture.

With reference to Figure 2, a gamma-densitometer 5 is added to the system of Figure 1 for periodic calibration.

The output of the gamma-densitometer 5 must be integregatec over several seconds to give an accurate result. In some flow regimes the gas and water fractions and can vary rapidly, and as the relation between these fractions and the mean permittivit and mean density are non-linear, averaging the two measurements independently before calculating the fractions can lead to errors. In this implementation (active reference) the fractions are calculated from the impedance measurement by a first compute unit 4, and the averaged fractions compared with fractions as calculated in the above described process, calculated by a secon computer unit 6. Any difference detected by the comparator 7 is used to calibrate the impedance measurement circuits, thus improving the (instantaneous) accuracy of the impedance.

With reference to Figure 3, the sensor comprises two electrodes 11 and 12 connected to terminals 13 and 14 respectively and having an electrode opening angle of 60°.

A guard 15 extends concentric with the outer screen and has an extension of approximately 180° of the circumference and assymetric to the associated electrode A. Said guard 15 is drive: at the same potential as that of the associated electrode A. The purpose of this guard is to reduce the stray capacitance from th electrode A to earth, i.e. to the sensor housing, to a minimum.

The most important parts of the sensor are the sensing electrode 11 and the live electrode 12. In principle, these electrodes could be insulated from the flow in the central cavit 18, however, any insulating layer would decrease the sensitivity to changes in the dielectrical properties of the flow. This is particularly important when measuring the dielectric loss, and non-insulated electrodes have therefore been chosen.

The electrodes 11 and 12 are each covering 120° of the circumference and thus leaves an opening covering an angle of 60 between the electrodes at either sides thereof. This opening angle has in the illustrated embodiment been chosen as a compromise between high sensitivity and homogenity of the electrical field in the sensor. Computer simulations have suggested that optimum choice of the electrode opening angle is 60° to 90°.

The electrode pair 11 and 12 will have to be shielded from external electrical fields and other sources of electrical noice and they have therefore been installed in a metal container whic also acts as sensor housing 16. The sensor housing is also completed with flanges to enable the sensor to be mounted flush with the associated flowline. The sensor housing is mounted concentrically around the electrodes and is separated from the electrodes by an electrically insulating material.

The separation between the electrodes 11, 12 and guard 15, and between the guard 15 and the sensor housing 16 has been chosen as a compromise between keeping a low capacitive load on the driver circuit, versus a practical overall size of the sensor.

The electrodes as illustrated have been constructed from stainless steel, but alternatively the electrodes may be constructed from other material with sufficient electrical conductivity and erosion resistance, such as an electrically conducting ceramic. The screen and the guard may be made from steel.

The space between the electrodes and the screen is filled with polyurethane insulation 17 leaving a central cavity 18 for fluid flow.

The polyurethane insulation has relatively low loss, a low permittivity and low water absorption. However, it is not resistant to erosion. In circumstances where this might be a problem, e.g. a production riser, it may be replaced by a wear-resistant insulating material such as a ceramic.

With reference to Figure 4, the sensor head equivalent curcuit is shown in therein. It consists of the bulk capacitance C_b shunted with the bulk loss conductance G_b, connected between nodes A and B.

Due to stray capacitances these nodes are connected to a common node G by C_A and C_B respectively. Each node A, B and G also have stray capacitances to earth, i.e. C'_A C'_B and C_G. Each capacitance is generally shunted by a conductance. These conductances are omitted as they carry a much smaller current compared to that of the accompanying capacitor.

The value of each component has been measured at +23°C with the sensor head empty. When measuring one component, all others have been guarded by the measurement bridge. The measurement results are given below.

Measured capacitances.

C_A = 43.50 pF	C'_A = 0.78 pF
C_B = 6.55 pF	C'_B = 21.32 pF
C_G = 104.60 pF	C_b = 3.06 pF

The small value of C'_A compared to C'_B is due to an inner co-axial semi-circular screen guarding electrode A in the sensor head.

As some of the electric stray fields also penetrate the bulk, the stray capacitances depend on the bulk mixture. Taking C'_B as an example of this dependency, it would be given as:

$C'_B = (21.3 \pm 2)$pF

If the stray capacitances had a fixed value, the impedance measured between nodes A and B in Figure 4 would only depend on bulk impedance G_b and C_b. But as this is not the case, the effects of stray capacitances have to be suppressed to an acceptable level.

With reference to Figures 5 and 6.

There are two ways in which the effects of stray capacitances (or any other impedance) may be eliminated.

Method 1: By ensuring that no driving voltage potential remains across the impedance. It then carries no current and may equally be removed from the circuit.

Method 2: By connecting the impedance across a source generator with zero output impedance. The voltage across the impedance will then be determined by the source generator and not by the impedance value or its fluctuations.

As it is impossible to ensure zero driving voltage potential, or zero output impedance, these methods will only suppress the effects of the stray capacitances.

The means of employing these methods are the circuits shown in Figures 5 - 6. In these circuits, the capacitances C_A, C_B and the input capacitance C_1 of the operational amplifier may or may not include the capacitances C'_A and C'_B; this depends on the circuit in question.

The open-loop gain of the op-amp is given by

$$A=A_o/(1+\tau s) \qquad [6]$$

where $s=\alpha+jw$ is the Laplacian variable.

The two circuits shown in Figures 5 and 6 are based on the concept that the current flowing through the conductance G_o is the same as that flowing through the parallel connection of G_b and C_b, i.e. the sensor.

If there is a conducting path from the sensor bulk volume to earth, i.e. the pipeline extending from both sides of the sensor, and if this path has a conductance not negligible compared to G_o, then the currents are no longer equal. The circuits will then have erroneous outputs.

Conductive paths will always exist as a result of axial stray capacitances, i.e. capacitances from electrode A or B to the extending pipelines. The value of these capacitances will vary with the bulk mixture. Calculations indicate variations less than $\pm$ 0.7pF. In addition, rather highly conductive paths occur when the water content of the bulk is above ~ 70%.

The bulk impedance components G_b and C_b may be determined by wideband or tuned circuits.

A wideband circuit, however, places great demands on phase measurements.

To avoid a system based on phase measurements, it was decided to investigate the possibilities of using a tuned circuit. By this means one would, instead of phase, measure the resonance frequency, which can be done to a far higher degree of accuracy.

Tuning the sensor by a shunting coil and driving the system at resonance, the current supplied to the sensor would only be that required by the sensor and shunt losses. This makes the sensor loss measurements much more accurate, even when they include the coil losses, compared to the situation, as in the

case of the wide-band circuits, where the loss current is overshadowed by a much larger reactive current supplied to the bulk capacitance.

Either of the two circuits shown in Figures 5 and 6 may be tuned. The 'driven-sensor' circuit of Figure 6 was selected as it provides both a peak and gain at resonance.

The resonance is:

$$\omega_p = 1/\sqrt{LC_T} \qquad [7]$$

where C_T is the toal capacitance between the sensor electrodes including shunting caused by the circuit, the coil, or capacitance for the purpose of stabilizing the amplifier.

The output voltage with an ideal amplifier is:

$$U_B = U_o \cdot \frac{G_o}{G_T} \qquad [8]$$

where G_T is the total conductance between the sensor electrodes including that of the tuning coil.

Letting $C_T = C_b + C_S$ [9]

and $G_T = G_b + G_L$ [10]

where b indicates the bulk components and S and L stand for shunt and coils components. This gives the following equations by which the bulk components may be calculated:

$$C_b = \frac{1}{\omega_p^2 L} \div C_s \qquad [11]$$

$$G_b = \frac{U_o}{U_B} G_o \div G_L \qquad [12]$$

As can be seen, C_b is determined solely by ω, and G_b solely by U_B, when the other circuit component values are known. In the real circuit, the relations are not quite so simple as above.

A change in G_b would move the resonance peak up or down in amplitude, i.e. a vertical movement. A change in C_b would move the resonance peak up or down in frequency, i.e. a horizontal

movement. Due to the nearly hyperbolic relation between C_b and R_b, the resonance peak will have low amplitude at low frequency and high amplitude at higher frequencies.

With reference to Figures 7 - 10

Figure 7 shows the equivalent tuned circuit of the 'driven-sensor' with no guard shown in Figure 5.

The operational amplifier has an output resistance R_G and an open loop gain given by equation 8 where A_o is the important dc-gain, and is the time constant. The gain cross-over frequency is $\omega_c = A_o/\tau$

The sensor and other components parallel to it, are represented by C_A, C_B, C_T and L, where C_T and G_T are defined by equations (9) and (10).

The op-amp output resistance R_G may be short-circuited, which also allows the capacitance C_B to be omitted.

The simplified equivalent circuit is thus as shown in Figure 8.

On the basis of Figure 8, the block diagram shown in Figure 9 may be drawn.

This may be re-arranged to achieve the ordinary feed-back loop shown in Figure 10.

Inserting the op-amp's open loop gain from equation (6), the forward gain of Figure 10 will be:

$$G(s) = \frac{1}{1+R_o C_A s} (-A) = \frac{-A_o}{(1+R_o C_A s)(1+\tau s)} \qquad [14]$$

and the feedback gain:

$$H(s) = \div(1+\frac{1}{A}) R_o Y_T = \div(1+\frac{1+\tau s}{A_o}) R_o (G_T + C_T s + \frac{1}{Ls}) \qquad [15]$$

The transfer function is thus:

$$\frac{U_B}{U_O} = \frac{G(s)}{1+GH(s)} = \frac{-LG_O s}{a_3 s^3 + a_2 s^2 + a_1 s + a_O} \qquad [16]$$

where, when assuming $A_O >> 1$, the coefficients are:

$$a_3 = \frac{L}{\omega_C}(C_A + C_T)$$

$$a_2 = L(C_T + \frac{C_A}{A_O}) + \frac{L}{\omega_C}(G_O + G_T)$$

$$a_1 = \frac{1}{\omega_C} + L(G_T + \frac{G_O}{A_O}) \qquad [17]$$

$$a_O = 1$$

Further simplifications may be done on the coefficients assuming the following magnitude relations to hold:

$$C_T >> \frac{C_A}{A_O} + \frac{1}{\omega_C}(G_O + G_T) \qquad [18]$$

$$G_T >> \frac{G_O}{A_O}$$

If so, the transfer function may be written as:

$$\frac{U_B}{U_O} = \frac{-LG_O s}{\frac{L}{\omega_C}(C_A + C_T)s^3 + LC_T s^2 + (LG_T + \frac{1}{\omega_C})s + 1} \qquad [19]$$

Considering the coefficients of the numerator in equation (19) it may be shown that the frequency of resonance is still given by equation (7) as long as:

$$\frac{C_A}{C_T} + 1 << \frac{\omega_C}{\omega_P} \qquad [20]$$

i.e. the bulk capacitance may be calculated when the frequency is measured and the permanent shunting capacitance C_S is known.

Thus, at the resonance peak the Laplacian variable has the value:

$$s_P = j\,\omega_P = j\frac{1}{\sqrt{LC_T}} \qquad [21]$$

When inserted into equation (19), the transfer function reduces to:

$$\frac{U_B}{U_o} = \frac{G_o}{G_T - \frac{\omega_p^2}{\omega_c} C_A} \qquad [22]$$

Solving this equation for the total conductance G_T and inserting from equation (10), gives the bulk conductance:

$$G_b = \frac{U_o}{U_B} G_o - G_L + \frac{\omega_p^2}{\omega_c} C_A \qquad [23]$$

Comparing this equation with (12), shows both the effect of the op-amp, by means of the gain cross-over frequency, ω , and the capactive loading C_A on the summing junction, i.e. the inverting input.

To calculate the bulk conductance G_b we must measure the reciprocal gain, i.e. U_o/U_B which is no major problem. Furthermore, the coil losses at the frequency in question have to be known. These losses may be taken from a look-up table measured at resonance with a special measurement system involving the coil in question and a high quality 'air-capacitor.'

Then, last but not least, we need to know the value of the stray capacitance C_A. As mentioned earlier, this capacitance has no fixed value, due to the fraction of the stray field penetrating the bulk.

As long as the three components of the bulk are thoroughly mixed, C_A may be expressed as a function of the bulk capacitance C_b, or as a function of the resonant frequency.

The actual relation between C_A and C_b has to be determined through measurements of these capacitances with different well-mixed contents of the sensor.

Shunting of the sensor

The conductive shunting

In order to prevent saturation at low sensor losses, one has to reduce the gain variations in the circuit as the bulk loss conductance varies. This is done by means of a conductance shunting the sensor. This conductance should not be too great, as it then tends to overshadow the bulk conductance.

The capacitive shunting

Disregarding the parallel sensor tuning, performed by the capacitance C_T and the coil L, there still exists a tuned loop in the electronics. This secondary loop comprises the operational amplifier, the total sensor losses G_T and the capacitive loading C_A on the summing junction.

The resonance frequency ω_n of this loop is given as:

$$\omega_c : \omega_n = \omega_n : \omega_A \qquad [24]$$

where

$$\omega_A = \frac{G_T}{C_A} \qquad [25]$$

In order to obtain an one-to-one correspondance between the sensor losses and the tuned curcuit output voltage level, one has to ensure that:

$$\omega_p < \omega_n \qquad [26]$$

The secondary loop determines the upper frequency limit for the sensor tuning.

If the sensor tuning ω_p is too close to the secondary tuning ω_n, variations in the peak at ω_n will influence on the voltage measured at ω_p, i.e. the output voltage.

This influence may be reduced by keeping the two frequencies sufficiently far apart.

This may only be achieved without any sacrifice by increasing ω_n which may be done by:

reducing the capacitive load on the summing junction, i.e. C_A, and increasing the gain cross-over frequency ω_c of the amplifier.

The usual means of achieving this is capacitance C_s shunting the feedback, i.e. the sensor.

To have the relative attenuation close to $\frac{1}{\sqrt{2}}$ (no peaking), the net shunting capacitance should have the value:

$$C_s' = 2\sqrt{2}\,\frac{G_T}{\omega_n} - \frac{G_o}{\omega_c} \qquad [27]$$

In our case, the net shunting capacitance will not have a fixed value with frequency, due to the neutralizing effect of the also shunting inductance. The net shunting is:

$$C_s' = (C_x + C_s)\ (\ 1 - [\frac{\omega_p}{\omega_n}]^2 \qquad [28]$$

Increased capacitive shunting also has the effect of reducing ω_p, which according to equation (26) is favourable. What is gained by reduction in ω_p, however, does not compensate for the loss in sensitivity and increased calibration uncertainty imposed by the reduced value of G_x.

<u>Estimate of accuracy obtainable</u>

Assume a system with a total sensor capacitance of 130pF and a resonance frequency at 440 KHz. When measuring this frequency with a resolution of $\pm$50Hz, the capacitance error will be:

$$dC_x = 2C_T\,\frac{d\omega_p}{\omega_p} = 2\,\frac{130 \cdot 10^{-12}}{2\,\pi\,440 \cdot 10^{2}}\,2\,\pi\,50 = 29{,}5 \cdot 10^{-15}F \qquad [29]$$

If C_b is 10% of the total shunting capacitance C_T, then the accuracy is:

$$100\%\ \frac{29{,}5 \cdot 10^{-15}}{0{,}1 \cdot 130 \cdot 10^{-12}} = \underline{0.23\%} \qquad [30]$$

Assuming further an imput conductance of 10uS, a 50pF loading on the summing junction and a gain cross-over frequency of 100 MHz; then, with gain of 20 and 0.2% voltage measurement accuracy together with 1nS uncertainty in the coil losses and a $\pm$ 0.2pF uncertainty in the value of C_A, the bulk conductance error will be:

$$dG_x = \frac{U_o}{U_B}\,G_o\,(\frac{dU_o}{U_o} + \frac{dU_B}{U_B}) + dG_L + C_A\,\frac{\omega_p^2}{\omega_c}\,(2\,\frac{d\omega_p}{\omega_p} + \frac{dC_A}{C_A}) \qquad [31]$$

$$= \frac{1}{20}\, 10\cdot 10^{-6}\ \cdot 2\frac{0,2}{100} + 1\cdot 10^{-9} + 50\cdot 10^{-12}\cdot 2\pi\,\frac{(440\cdot 10^{3})^{2}}{100\cdot 10^{6}}\left(2\frac{50}{440\cdot 10^{2}}+\frac{0,2}{50}\right)$$

$$\approx 5,6\cdot 10^{-9}\ \text{Siemens}$$

With a bulk conductance of 100 nS (10M) the accuracy will be:

$$100\%\ \frac{5.6}{100} = 5.6\%$$

Conclusion regarding tuned circuits

The accuracy of the calculated bulk conductance G_x depends greatly on precise knowlegde of the stray capacitance C_A. This capacitance varies ± 10% about a fixed value, and can be described by the resonance frequency when the bulk is thoroughly mixed. Voltages have to be measured with 0.2% accuracy or better, which requires good long-term stability demodulators, and accurate calibration of the voltage measuring system.

The frequency measurements are the least critical, as a resolution of ±50Hz is sufficient. This could be achieved when measuring the total period of 100 cycles with a 50MHz oscillator as a time base, when the resonance frequency is about 500 KHz.

The circuit will have to be excited at the correct frequency ω_p, when making the frequency and voltage measurements. This could be done, either by swept measurements or by connecting the tuned amplifier as an oscillator. In the latter case the system will be 'self-tracking.' With reference to Figures 11 to 15,

The tuned amplifier comprises a operational amplifier, a coil, an input resistance and a shunting capacitance.

The parallel connection of R_4 and C_5 is inserted to shape the frequency response above 10MHz.

The operational amplifier is of the type DATEL-INTERSIL: AM-500MC ultra-fast, inverting only. Its equivalent circuit is shown in Figure 12.

The complete transfer function of the amplifier in Figure 12 is given as:

$$A(s) = \frac{10^{6}}{(1+s\ 1,59\cdot 10^{-3})(1+s\ 227\cdot 10^{-12})} \qquad [32]$$

It has been shown that it is sufficient to use only the low frequency time constant $\tau = 1.59.10^{-3}$ in equation (32) when doing calculations on the circuit. With the amplifier mounted on the printed curcuit card, the gain cross-over frequency is reduced to 76.7MHz. The low frequency time constant thus has to be increased to $\tau = 2.077.10^{-3}$ sec.

Measurements have shown that when amplifier temperature increase its low-frequency time contant τ is further increased, i.e. the phase margin of the tuned amplifier is reduced and the system becomes more oscillatory.

The coil

The coil consists of 120 turns of stranded wire wound on a 3-chamber coil-former enclosed in a pot-core. The wire has 15 strands, each of 0,05mm diameter. The pot-core is a Philips P18/11 with core material 3D3. Its effective permeability is u_e=33. The coil was designed to have an inductance of L = 1mH at a frequency of 1MHz, and its quality factor Q should be greater than 250. The resulting Q-factor was measured to 390, i.e. the coil's loss-conductance was 400 nS at the design frequency.

The shunting self-capacitance of the coil was calculated to 2.44pF, based on the self-resonance which was measured to 3.22117 MHz.

The coil was artifically aged by means of temperature cycling. Several measurements have shown that the coil needs to be continuosly energized for a minimum of 18 hours before its losses have stabilized to within +1nS. The coil has to be temperature stabilized to within a few degrees, and it should also be in a dry nitrogen atmosphere to exclude losses caused by humidity. When stabilized, the coil loss conductance may be expressed as a non-linear function of frequency.

The input conductance, G_o

This input conductance G_o has the value 12.223 S. There is no observable change in this value when the frequency is varied between 400 KHz and 600 KHz.

The conductance is shunted by its own stray capacitance and the self-capacitance of the circuit layout. These add up to a capacitance of 0.17pF which is frequency independent. The

shunting effect of this capacitance may be neglected for frequencies below 1/10 of the half-power bandwidth of the low-pass filter formed, i.e. for frequencies below 1.2MHz.

<u>The shunting capacitance, C_S</u>

It consists of the self-capacitance of the coil (0.22pF) and the added fixed and variable capacitances. The total, when adjusted, is measured to 135.54pF. The maximum frequency with L=1mH is thus 432.3KHz.

<u>The capacitive input load, C_A</u>

The capacitive load on the inverting input consists of the following parts:

Sonsor stray capacitance $C_B + C'_B$	= 27.87	pF
Self-capacitance of co-ax	= 5.195	pF
Circuit layout capacitance	= 3.27	pF
Op-amp input inpacitance	= 2	pF
Coil windings to earth capacitance	= 11.715	pF

The total sum is 50.05 + 2pF, as it depends on the axial stray capacitances and the non-fixed value of C_B as given in equation (5).

<u>Bulk capacitance vs. resonance frequency</u>

Each resonance frequency is determined by swept measurements, using different values of fixed capacitors as substitutes for the bulk capacitance.

The resonance frequency is where the amplifier gain has a peak. The instrument used for this purpose is a network analyser type 3577A made by Hewlett Packard. This instrument has a peak-search capability.

Resonance frequency vs. bulk capacitance C_b is as shown in Figure 14 when C_s = 120.33pF

Measurements have been carried out to verify that the resonance frequency is independent of the bulk conductance.

Fixed resistors were used as substitutes for the bulk conductance, with different values C_b. These measurements confirmed the independence of ω_p and G_L.

As a good approximation to the relation between the resonance frequency (in KHz) and the bulk capacitance (in pF), one may use the following equation (also with C_s = 120.33 pF):

$$f_p = \frac{5081111.7}{\sqrt{C_b + 120.33}} \qquad [\,33\,]$$

The maximum rate of change in the bulk capacitance has been determined by means of the response in the resonance frequency when making an abrupt change in the bulk capacitance, i.e. its substitute C_b. The response of the resonance frequency was obtained by means of an FM-demodulator.

The following results were obtained:

at C_b = 4.6 pF max. rate 0.9 pF/ms
at C_b = 30 pF max. rate 18.5 pF/ms
at C_b = 77 pF max. rate 134 pF/ms

When the bulk capacitance has a low value, for example when the void fraction is significant, the expected capacitance change is also small.

The complete measuring circuit is shown in Figure 15.

When the tuned amplifier is working at its resonance frequency, it is purely an inverting amplifier which produces a phase-shift of 180° from input to output. In order to lock the amplifier at resonance, a phase-locked loop is connected to it. The phase-locked look consists of a phase comparator, a loop filter (active lowpass filter) and a voltage-controlled oscillator (VCO).

The 'phase-comparator' detects any change in frequency or phase between the input and the output of the amplifier. A phase-shift other than 180° will cause an imbalance between the signals U_1 and U_3 and force the 'loop filter' to change the VCO control voltage until the frequency of the VCO is exactly equal to the resonance frequency of the amplifier, i.e. the loop is locked. When the loop is unlocked, a 'lock indicator' takes control of the VCO and sweeps its frequency until it equals the resonance frequency of the amplifier, and the loop returns to lock.

The VCO output signal is a triangular waveform of constant amplitude. The frequency range is set within higher and lower limits, according to the actual working range of the sensor.

To calculate the gain of the amplifier (U/U_0), a precision voltmeter is used in conjunction with two fullwave rectifiers. Because the input signal U_0 is small, a buffer amplifier of fixed gain 26 dB is used prior to rectification.

The output frequency of the amplifier is tapped after zero detection in the phase comparator to avoid overloading of the amplifier output.

CLAIMS.

1. Process for determining the proportions of the components of a three-component mixture, especially the proportions of gas, water and oil in a gas/water/oil mixture, characterised in

measuring the permittivity and the dielectric loss in the three-component mixture during passing said mixture in a flow through a spacing between two opposite electrodes,

processing signals received from said electrodes in a computer by incorporating the following equations therein:

$$\frac{1-\varepsilon_b}{1-\varepsilon_s}\left(\frac{\varepsilon_s}{\varepsilon_b}\right)^{1/3} = 1-\alpha \qquad [1]$$

where α is a first component (gas), ε_b is the permittivity of the three-component mixture, ε_s is the permittivity of a two-component mixture of a second (water) and a third (oil) component and is given by:

$$\varepsilon_s = \frac{\varepsilon_o}{(1-\beta_f)^{3-4.2} \cdot \beta_f^{6} + 3\,\beta_f^{10}} \qquad [1a]$$

where β_f is the concentration of the second component in said two-compoent mixture and the actual fraction of the second component in the three-component mixture is given by:

$$\beta=(1-\alpha)\beta_f \qquad [1b]$$

and

$$\sigma_b = \frac{\sigma_s}{\varepsilon_s}\{\varepsilon_b - \alpha[\,\alpha + (\frac{3\,\varepsilon_s}{1+2\,\varepsilon_s})^2(1-\alpha)\,]\} \qquad [2]$$

where σ_s is the dielectric loss of said two-component mixture, ε_s is the permittivity of said two-compoent mixture, and ε_b is the permittivity of the three-component mixture and is theoretically given by

$$\sigma_s = \sigma_o\frac{\varepsilon_s}{\varepsilon_o} \qquad [2a]$$

and progressively determining the proportions of said three components in said three-component mixture.

2. Instrument for determining the proportions of the components in a three-component mixture, and especially the proportions of gas, water and oil in a gas/water/oil mixture, characterised in

a) an electrode impedance sensor located in a concentrated flow of said three-component mixture for measuring the permittivity and the dielectric loss of said three-component mixture,

b) an impedance sensor head for reducing or eliminating impedance between the electrodes and a screen,

c) impedance electronics comprising a tuned amplifier and a phase locked loop circuit forming a tracking resonance system, for maintaining the resonance frequency, and

d) a computer for processing the electronic signals and determining the proportions of said three components (gas, water and oil) in said three-component mixture.

3. Instrument according to claim 2, characterised in

incorporation of a gamma densitometer for periodic calibration of the instrument of claim 2.

4. Instrument according to claim 2 or 3, characterised in

that the electrodes of the sensor head are in the shape of arcuate electrode plates flushing with an inner surface of a tube forming sensor head.

5. Instrument according to one of claims 2 to 4, characterised in

that the radially innermost surface of the electrode plates are in direct, i.e. un-insolated contact with the mixture flow.

6. Instrument according to one of claims 4 to 5, characterised in

that the electrodes circumferencially along the inner surface of the tube forming sensor head are mutually spaced b opening angle between approximately 60 to 90°.

7. Instrument according to one of claims 2 to 4, c h a r · a c t e r i s e d i n

that the sensor head is ring shaped or sleeve shaped wi an radially outermost house forming screen and a pair of radi innermost electrodes and in a volume between same an electric insulating material.

8. Instrument according to claim 7, c h a r a c t e r i s i n

that the electrically insulating material is a ceramic o polyurethane.

9. Instrument according to one of claims 2 to 8, c h a r - a c t e r i s e d i n

that the electrode is made of an electrically conducting ceramic.

10. Instrument according to one of claims 7 to 9, c h a r - a c t e r i s e d i n

that the house forming screen is at earth potential,

that a guard is located between the screen and at least a first one of the electrodes, and

that said guard is at the potential of said first electrode.

11. Instrument according to claim 10, c h a r a c t e r i - s e d i n

that the guard has an arcuate extension of approxiamately 180° symmetrically of said first electrode.

12. Instrument according to claim 10, c h a r a c t e r i - s e d i n

that the guard has an arcuate extension of 360° concentric of the electrodes.

13. Instrument according to one of claims 7 to 9, c h a r a c t e r i s e d i n

that the house forming screen is at earth potential,

that the sensor head at the radial innermost surface in each spacing between the electrodes is provided with a strip shaped guard with an extension of some few arcuate degrees, and

that said guard strips are at the potential of said first electrode.

14. Instrument according to one of claims 2 to 13, c h a r a c t e r i s e d i n

that the impedanace electronics are thermostatically controlled.

AMENDED CLAIMS

[received by the International Bureau on 13 February 1990 (13.02.90) original claims 1 and 2 amended,other claims unchanged (2 pages)]

1. Process for determining the proportions of the components of a three-component mixture, especially the proportions of gas, water and oil in a gas/water/oil mixture, characterised in

measuring the permittivity and the dielectric loss in the three-component mixture by using non-intrusive measurement instrument during passing said mixture in a flow through a spacing between two opposite electrodes,

processing signals received from said electrodes in a computer by incorporating the following equations therein:

$$\frac{1-\varepsilon_b}{1-\varepsilon_s}\left(\frac{\varepsilon_s}{\varepsilon_b}\right)^{1/3} = 1-\alpha \qquad [1]$$

where α is a first component (gas), ε_b is the permittivity of the three-component mixture, ε_s is the permittivity of a two-component mixture of a second (water) and a third (oil) component and is given by:

$$\varepsilon_s = \frac{\varepsilon_o}{(1-\beta_f)^{3-4.2\cdot\beta_f^{6}+3\beta_f^{10}}} \qquad [1a]$$

where β_f is the concentration of the second component in said two-compoent mixture and the actual fraction of the second component in the three-component mixture is given by:

$$\beta = (1-\alpha)\beta_f \qquad [1b]$$

and

$$\sigma_b = \frac{\sigma_s}{\varepsilon_s}\{\varepsilon_b - \alpha[\alpha + (\frac{3\varepsilon_s}{1+2\varepsilon_s})^2(1-\alpha)]\} \qquad [2]$$

where τ_s is the dielectric loss of said two-component mixture, ε_s is the permittivity of said two-compoent mixture, and ε_b is the permittivity of the three-component mixture and is theoretically given by

$$\sigma_s = \sigma_o\frac{\varepsilon_s}{\varepsilon_o} \qquad [2a]$$

and progressively determining the proportions of said three components in said three-component mixture.

2. Instrument for determining the proportions of the components in a three-component mixture, and especially the proportions of gas, water and oil in a gas/water/oil mixture, c h a r a c t e r i s e d i n

a non-intrusive measurement instrument comprising

a) an electrode impedance sensor located in a concentrated flow of said three-component mixture for measuring the permittivity and the dielectric loss of said three-component mixture,

b) an impedance sensor head for reducing or eliminating impedance between the electrodes and a screen,

c) impedance electronics comprising a tuned amplifier and a phase locked loop circuit forming a tracking resonance system, for maintaining the resonance frequency, and

d) a computer for processing the electronic signals and determining the proportions of said three components (gas, water and oil) in said three-component mixture.

3. Instrument according to claim 2, c h a r a c t e r i s e d i n

incorporation of a gamma densitometer for periodic calibration of the instrument of claim 2.

4. Instrument according to claim 2 or 3, c h a r a c t e - r i s e d i n

that the electrodes of the sensor head are in the shape of arcuate electrode plates flushing with an inner surface of a tube forming sensor head.

5. Instrument according to one of claims 2 to 4, c h a r - a c t e r i s e d i n

that the radially innermost surface of the electrode plates are in direct, i.e. un-insolated contact with the mixture flow.

6. Instrument according to one of claims 4 to 5, c h a r - a c t e r i s e d i n

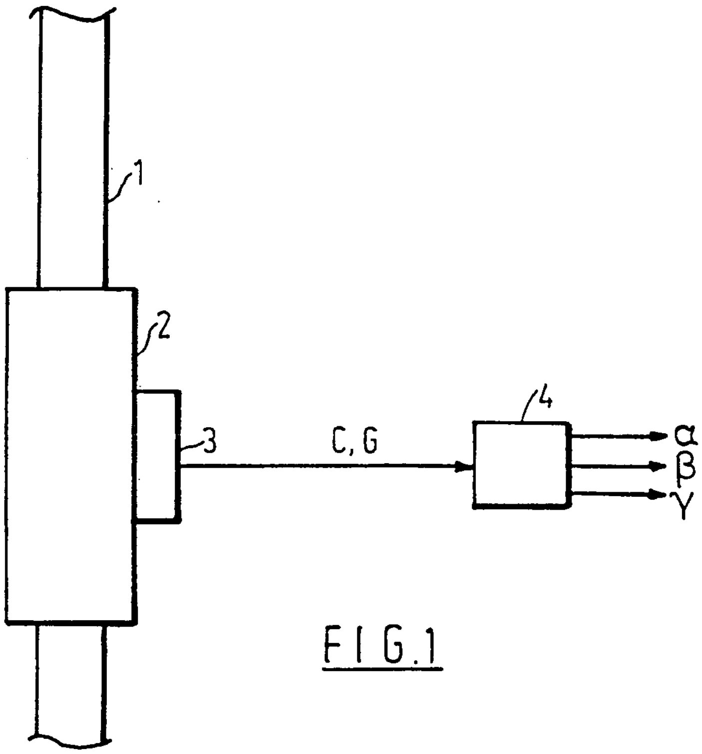


FIG.1

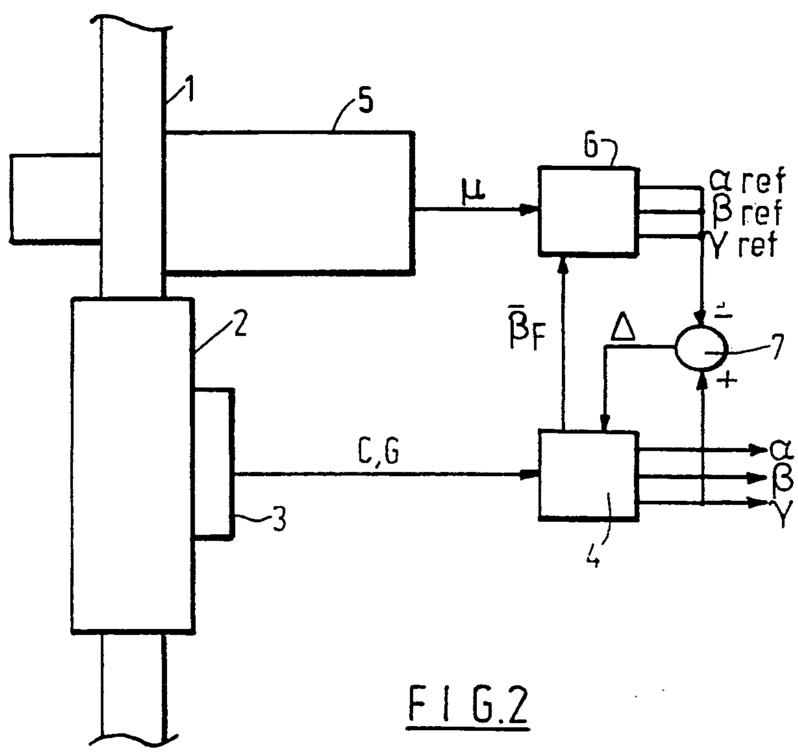


FIG.2

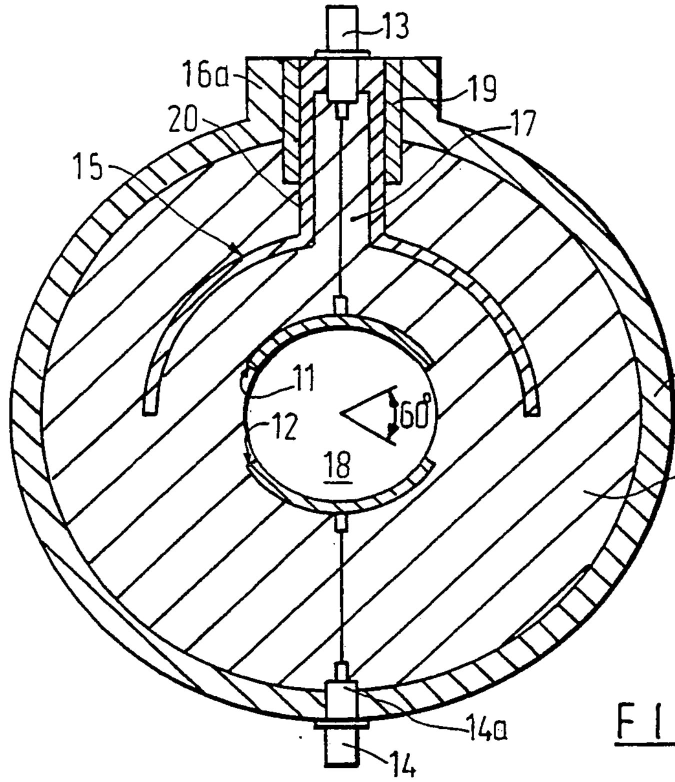

13
16a
19
20
17
15
11
60°
12
18
16
17
14a
14

FIG.3

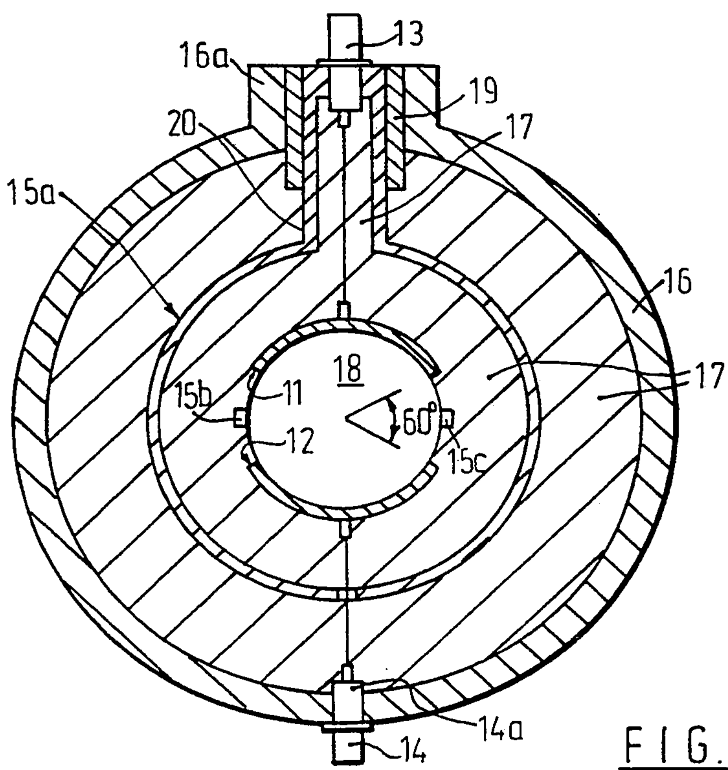

13
16a
19
20
17
15a
16
18
15b
11
60°
17
12
15c
14a
14

FIG.16

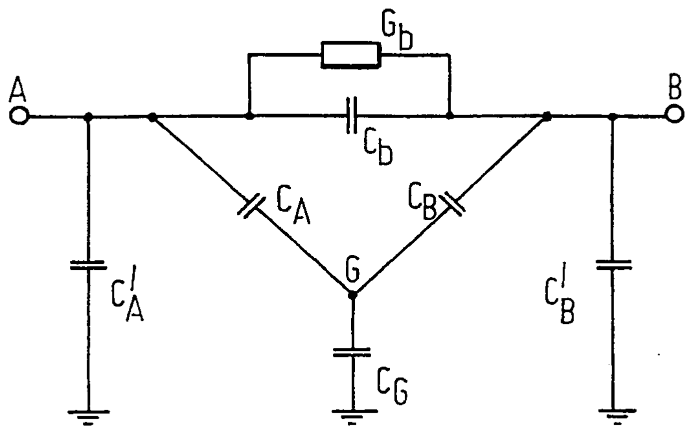


FIG.4

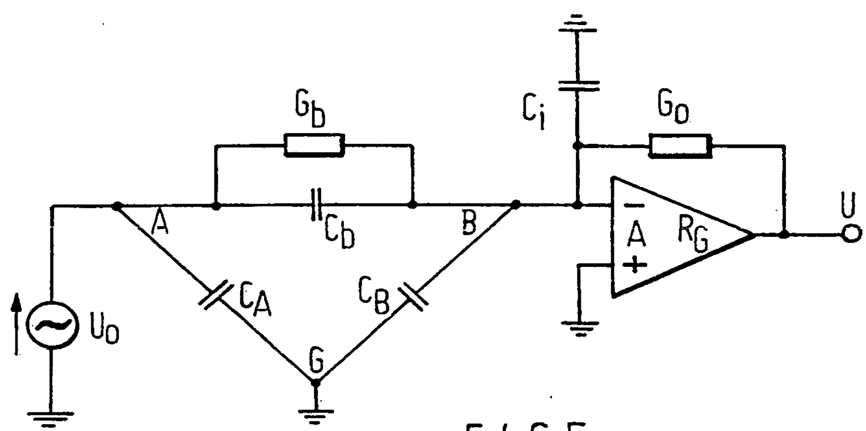


FIG.5

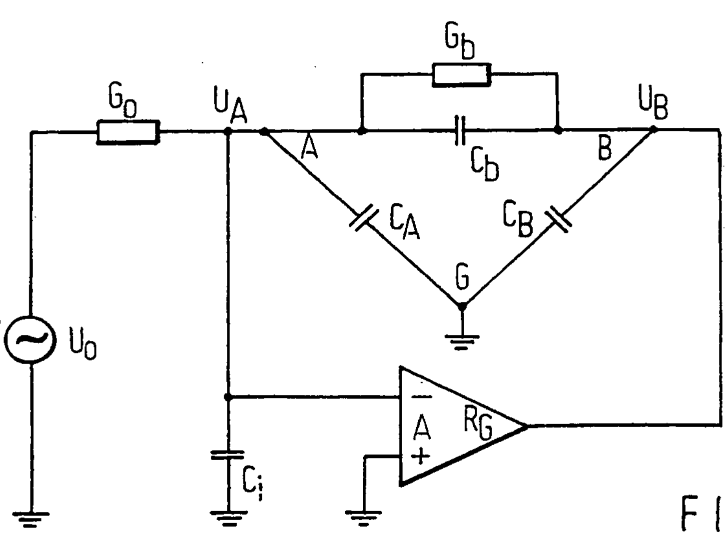


FIG.6

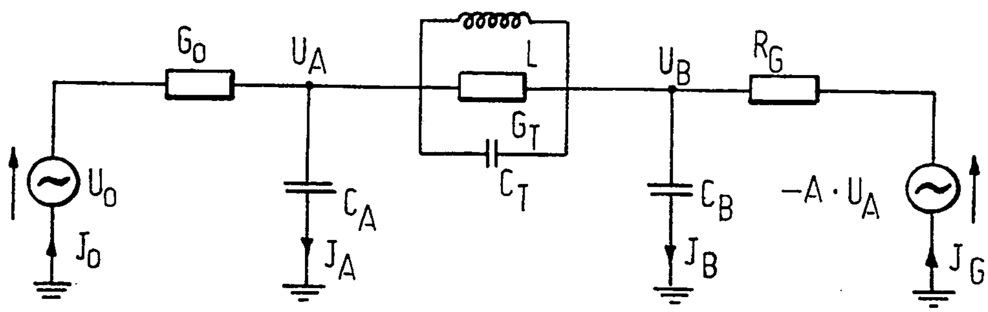


FIG.7

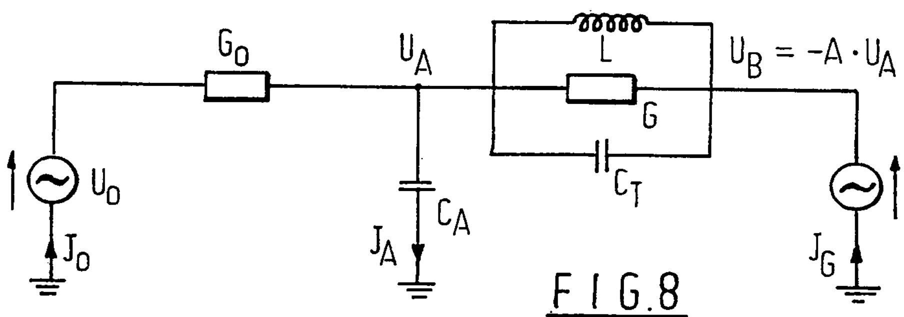


FIG.8

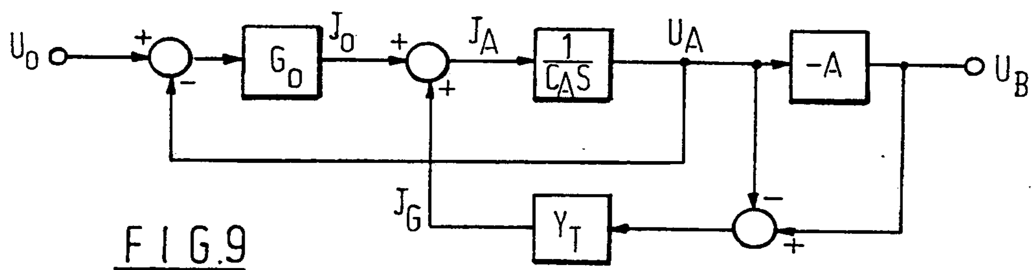


FIG.9

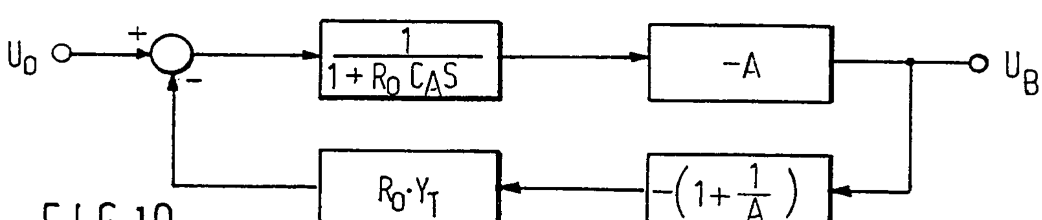
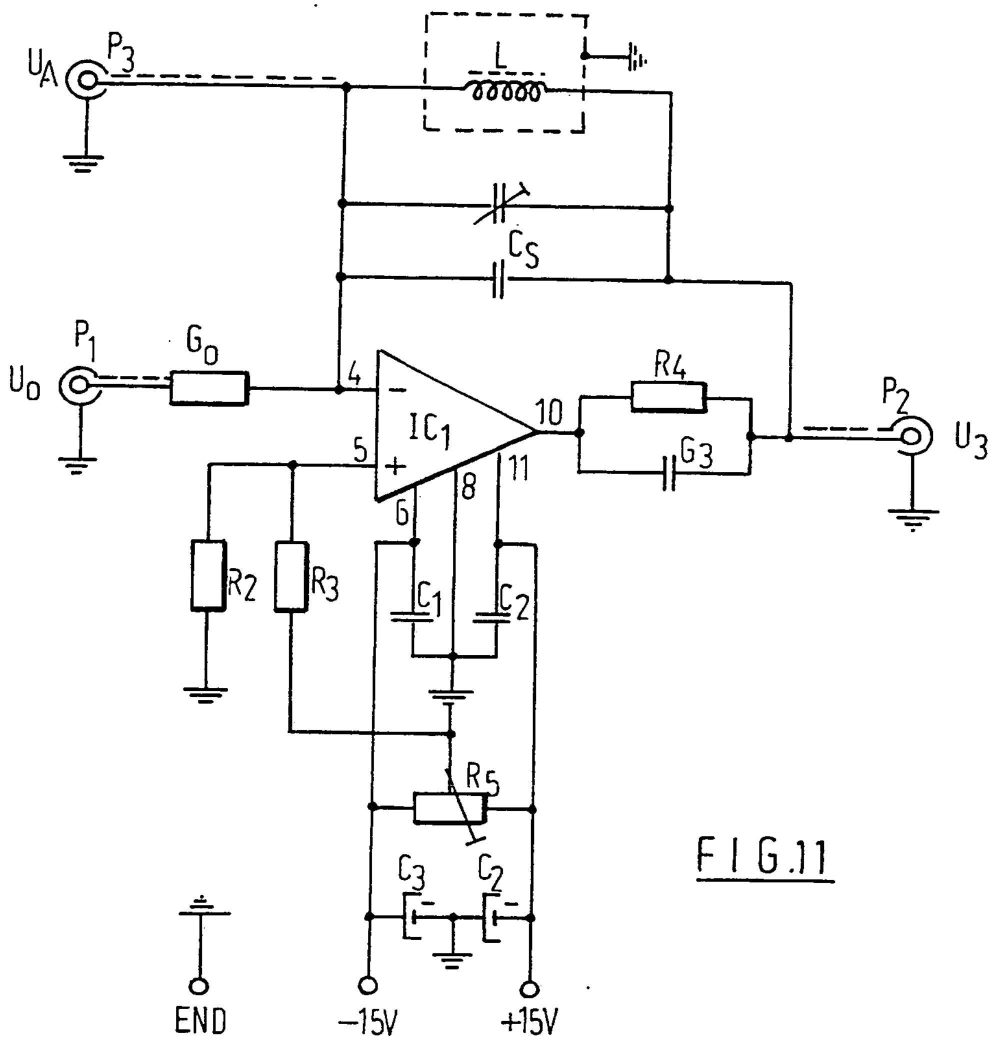
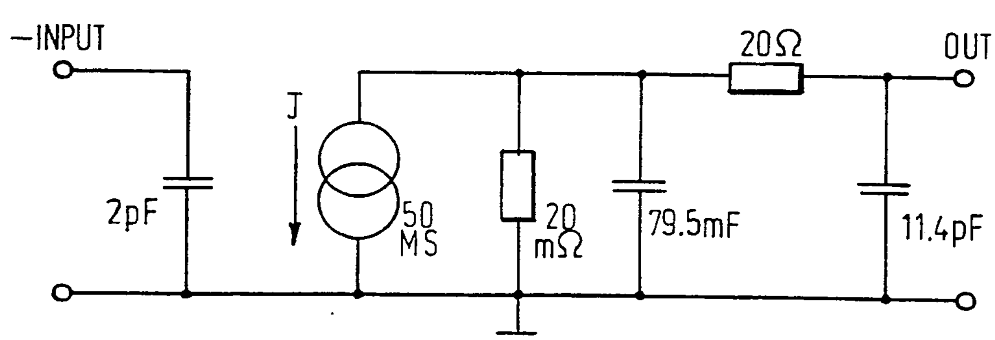


FIG.10



U_A
P_3
L
C_S
P_1
G_0
U_0
4
5
IC_1
10
R4
P_2
U_3
G_3
6
8
11
R2
R3
C_1
C_2
R_5
C_3
C_2
END
−15V
+15V

FIG.11



—INPUT
J
2pF
50 MS
20 mΩ
79.5mF
20Ω
OUT
11.4pF

FIG.12

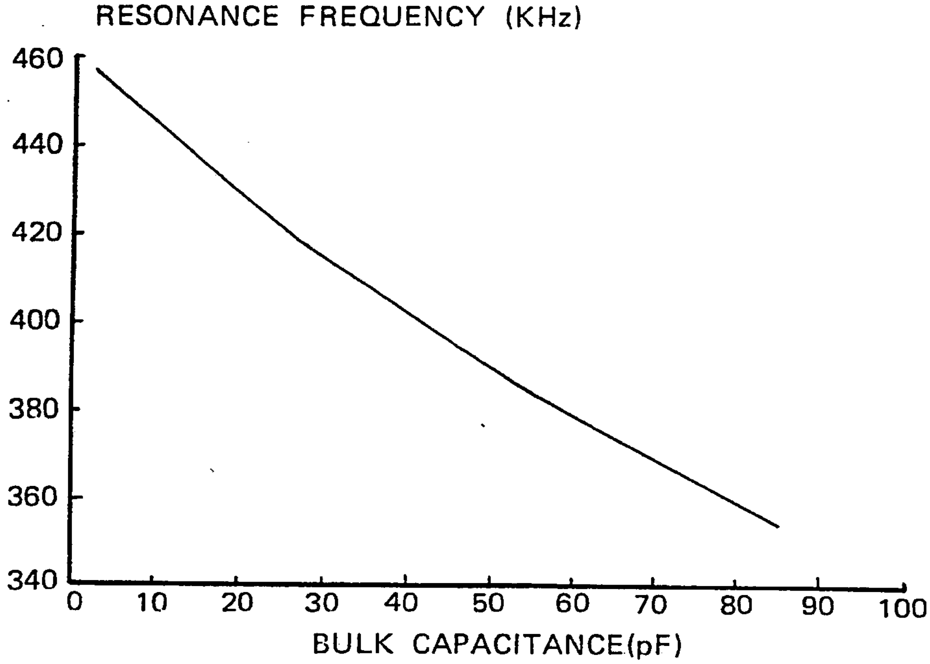


FIG.13

BULK LOSS CONDUCTANCE
v.s. GAIN AND BULK CAPACITANCE

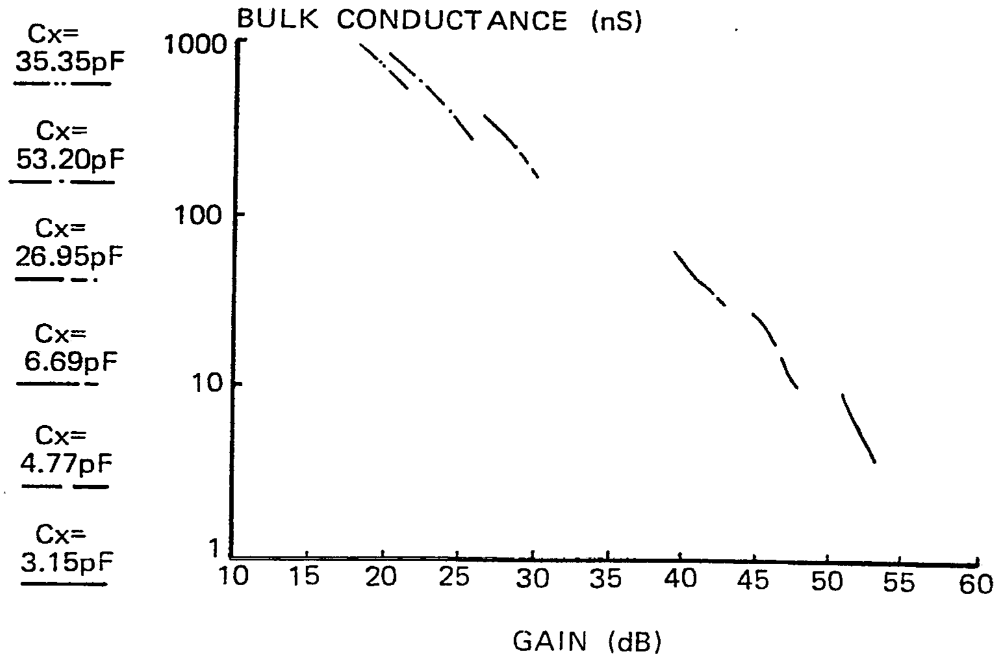


FIG.14

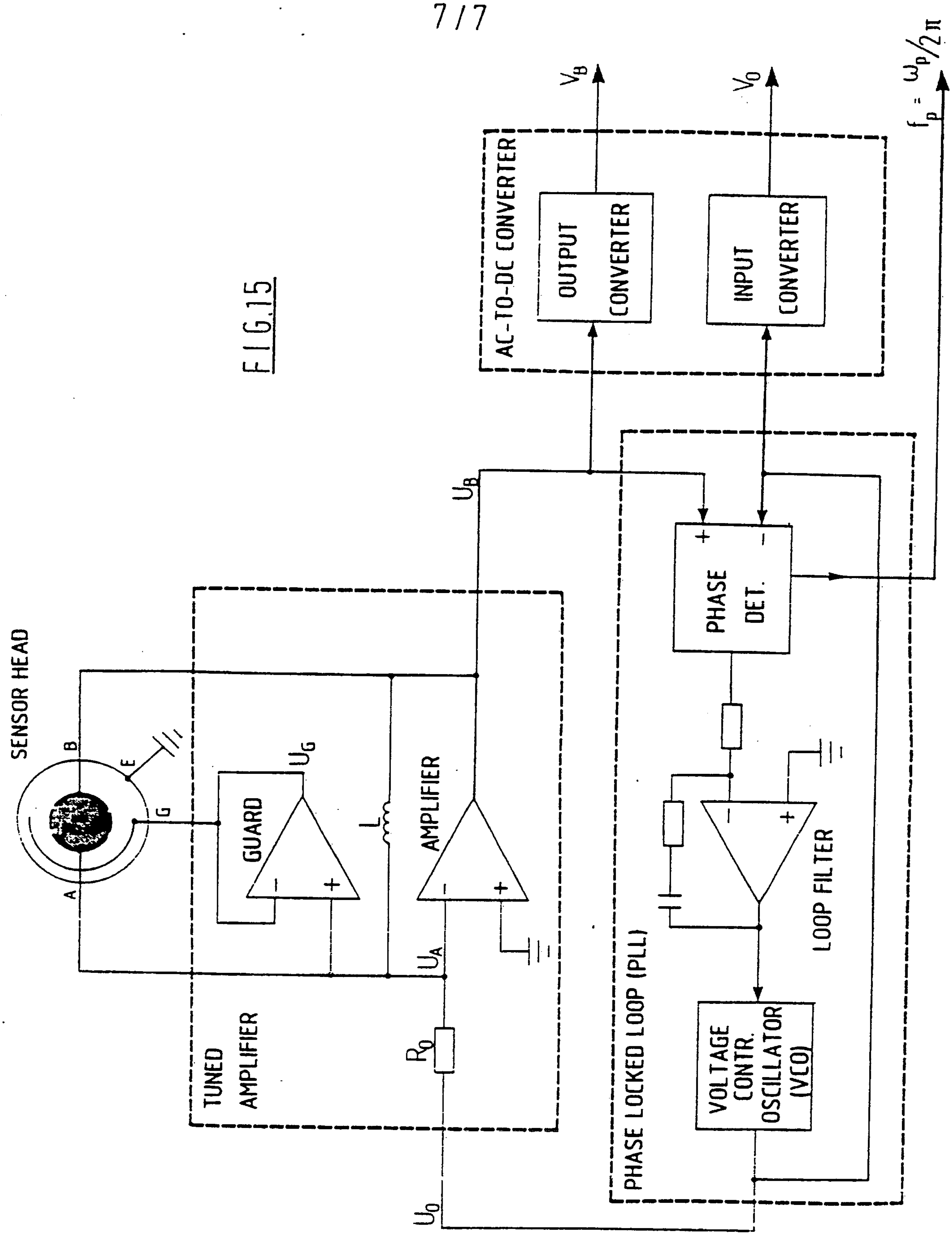

SENSOR HEAD
A
B
E
G
GUARD
U_G
L
AMPLIFIER
U_A
U_B
R_0
TUNED AMPLIFIER
U_0
AC-TO-DC CONVERTER
OUTPUT CONVERTER
INPUT CONVERTER
V_B
V_0
$f_p = \omega_p/2\pi$
PHASE DET.
LOOP FILTER
VOLTAGE CONTR. OSCILLATOR (VCO)
PHASE LOCKED LOOP (PLL)

FIG.15

INTERNATIONAL SEARCH REPORT

International Application No PCT/NO 89/00087

I. CLASSIFICATION OF SUBJECT MATTER (if several classification symbols apply, indicate all) [6]

According to International Patent Classification (IPC) or to both National Classification and IPC

IPC4: G 01 N 27/10, 23/08

II. FIELDS SEARCHED

Minimum Documentation Searched [7]

Classification System	Classification Symbols
IPC4	G 01 N

Documentation Searched other than Minimum Documentation to the Extent that such Documents are Included in the Fields Searched [8]

SE,DK,FI,NO classes as above

III. DOCUMENTS CONSIDERED TO BE RELEVANT [9]

Category *	Citation of Document, [11] with indication, where appropriate, of the relevant passages [12]	Relevant to Claim No. [13]
Y	J. Phys. E: Sci. Instrum., Vol. 18, 1985 (Great Britain) Eivind Dykesteen et al: "Non-intrusive three-component ratio measurement using an impedance sensor ", --	1,2
Y	US, A, 3675121 (DON D. THOMPSON) 4 July 1972, see the whole document --	1
Y	US, A, 4644263 (IRVIN D. JOHNSON) 17 February 1987, see the whole document --	2-3
Y	GB, A, 2088050 (ERNEST JOHN MICHAEL KENDALL) 3 June 1982, see the whole document --	3

* Special categories of cited documents: [10]

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IV. CERTIFICATION

Date of the Actual Completion of the International Search	Date of Mailing of this International Search Report
5th December 1989	1989 -12- 11
International Searching Authority	**Signature of Authorized Officer**
SWEDISH PATENT OFFICE	Eva Iversen Hasselrot

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International Application No. PCT/NO 89/00087

III. DOCUMENTS CONSIDERED TO BE RELEVANT (CONTINUED FROM THE SECOND SHEET)		
Category *	Citation of Document, with indication, where appropriate, of the relevant passages	Relevant to Claim No
A	US, A, 4713603 (RICHARD THORN) 15 December 1987, see the whole document --	1-14
A	DE, A, 1798370 (ALCO STANDARD CORP.) 20 January 1972, see the whole document --	2-14
A	DE, C2, 2533943 (GEORGE KENT LTD.) 21 March 1985, see the whole document --	2-14

ANNEX TO THE INTERNATIONAL SEARCH REPORT
ON INTERNATIONAL PATENT APPLICATION NO. PCT/NO 89/00087

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GB-A- 2088050	03/06/82	JP-A- 57116240	20/07/82
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US-A- 4713603	15/12/87	WO-A- 85/02016	09/05/85
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